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ROBUST INTERFERENCE MITIGATION AND SPREAD SPECTRUM SIGNALING

State University of New York at Buffalo

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APPROVED: /s/

JOHN D. MATYJAS
Project Engineer

FOR THE DIRECTOR: /s/

WARREN H. DEBANY JR., Technical Advisor
Information Grid Division
Information Directorate

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6. AUTHOR(S) Lisimachos P. Kondi				
7. PERFORMING ORGANIZATION NAME(S) AND ADDRESS(ES) State University of New York at Buffalo Dept of Electrical Engineering 332 Bonner Hall Buffalo New York 14620			8. PERFORMING ORGANIZATION REPORT NUMBER N/A	
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Table of Contents

Introduction	1
Background	1
Results	4
C Implementation of the Developed Algorithms	8
Publications	9

List of Appendixes

Appendix A - Interference Resistant Scalable Video Transmission Over DS-CDMA Channels	10
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List of Figures

Figure 1: Flow Diagram	2
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List of Tables

Table 1: Results for 120kbps clip (142 frames) No Doppler for all users	4
Table 2: Results for 40kbps clip (300 frames) No Doppler for all users	5
Table 3: Results for 40 kbps clip (300 frames) 4 Hz Doppler for User of Interest; No Doppler for Interferers	5
Table 4: Results for 40 kbps clip (300 frames) with 40 Hz Doppler 40 Hz Doppler for User of Interest; No Doppler for Interferers	6
Table 5: Results for 40 kbps clip (300 frames) 200 Hz Doppler for User of Interest; No Doppler for Interferers	6
Table 6: Results for 40 kbps clip (142 frames) 4 Hz Doppler for all users	7
Table 7: Results for 40 kbps clip (142 frames) 40 Hz Doppler for all users	8
Table 8: Results for 40 kbps clip (142 frames) 200 Hz Doppler for all users	8

Introduction

Transmission of direct-sequence code-division multiple access (DS-CDMA) video data streams over an actual RF channel environment with Rayleigh fading is studied to validate the superiority of auxiliary-vector (AV) filters over other filters. The AV filter is used at the receiver since it is known to outperform other filtering techniques in rapidly changing environments, where only a small data record size is available.

Background

The flow diagram of the process used in this study is shown in Figure 1. We used the Agilent E4438C RF Arbitrary Waveform Generator, the N115A Baseband Studio fader hardware/software package, and the SoRDS receiver to provide an actual RF channel environment with Rayleigh fading. The data generated included data streams for 10 users and Doppler values equal to 0, 4, 40, 200, and 400 Hz.

The transmitted video data was the "Foreman" sequence source-encoded using MPEG-4. The video data stream is partitioned into small packets. Each packet is then Rate-Compatible Punctured Convolutional (RCPC) encoded, interleaved, and then spread.

Convolutional coding is accomplished by convolving the source data with a convolutional matrix G , thus generating one codeword for the entire source data. RCPC codes are well-suited for unequal error protection since they provide a family of codes with different rates. Rate-compatibility requires that a higher rate code be a subset of a lower rate code. This is achieved by puncturing a "mother" code of rate $1/n$ to achieve high rates. Puncturing is the process of deleting bits from the output sequence in a predefined manner so that fewer bits are transmitted than in the original coder leading to a higher coding rate.

Using interleaving, as well as an error correcting code, can improve the performance of channel coding. Interleaving changes the mixes symbols from different codewords so that consecutive symbols within the same codeword are spread across several codewords. Error bursts would then affect different symbols belonging to different codewords instead of a whole section of one codeword. This has the effect of randomizing burst errors seen by the channel decoder.

DS-CDMA allows several users to use the same frequency band at the same time. Each user is assigned a unique code called the spreading code that is used to distinguish one user from the other. The spreading length used in this investigation is 16.

As a result of the limitations of the hardware used to create the RF channel environment, only one user could be transmitted at a time. To provide the more realistic multiuser scenario, individual user data was summed at the receiver.

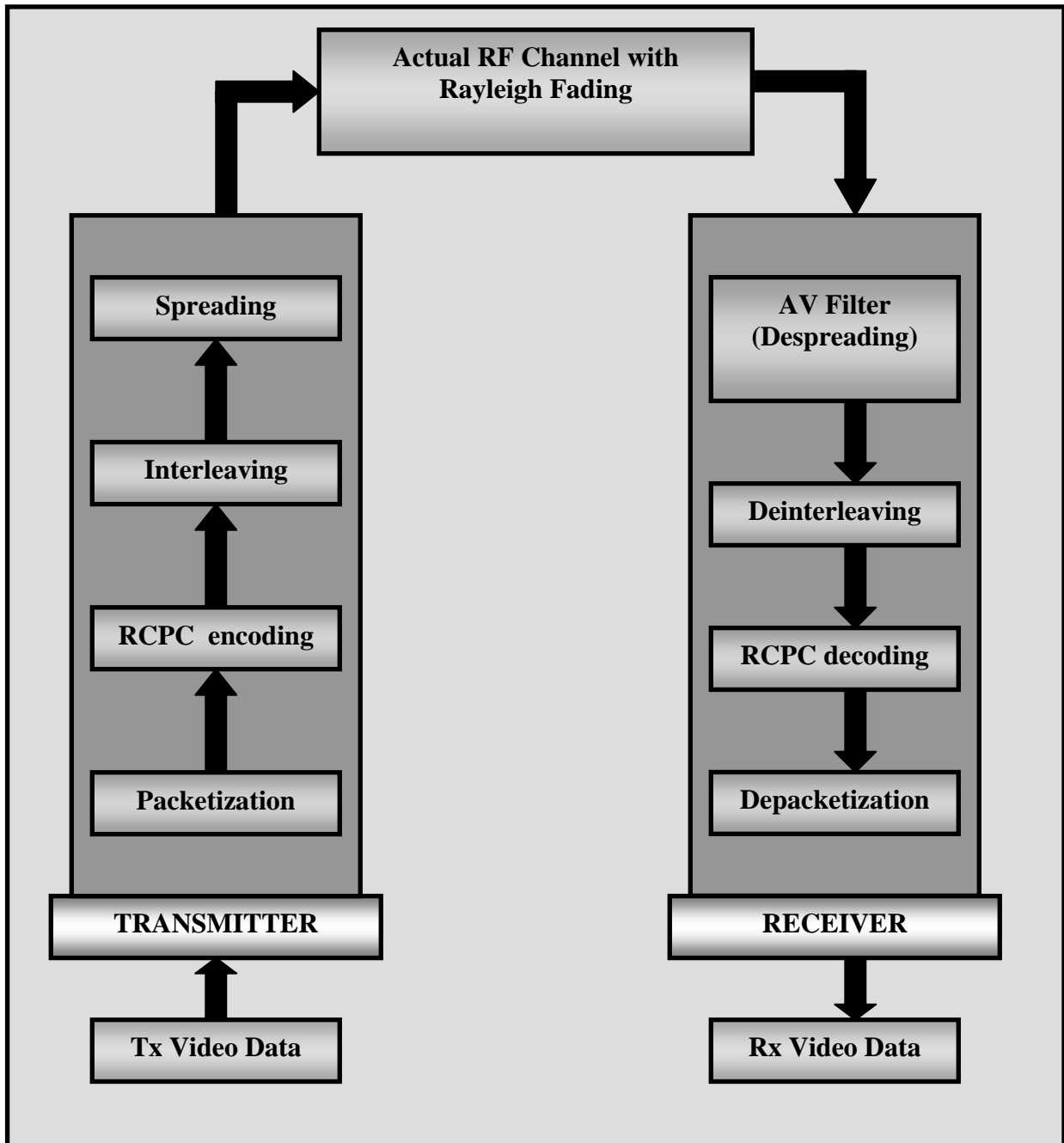


Figure 1: Flow Diagram

The AV receiver was chosen based on the realistic channel fading rates that limit the data record available for receiver adaptation. Under small sample support adaptation, AV filter short-data-record estimators have been shown to exhibit superior bit error rate (BER) performance in comparison to least mean squares (LMS), recursive least squares (RLS), sample matrix inversion (SMI), diagonally loaded SMI, or multistage nested Wiener filter implementations.

The AV algorithm generates an infinite sequence of filters $\{w_k\}_{k=0}^{\infty}$. The sequence is initialized to the space-time RAKE matched filter (MF):

$$w_0 = \frac{w_{R-MF}}{\|w_{R-MF}\|^2} \quad (1)$$

which is scaled to satisfy $w_0^H w_{R-MF} = 1$. At each step $k+1$ of the algorithm, $k=0,1,2,\dots$, etc. an "auxiliary" vector component g_{k+1} that is orthogonal to w_{R-MF} is incorporated in w_k and weighted by a scalar μ_{k+1} to form the next filter in the sequence:

$$w_{k+1} = w_k - \mu_{k+1} g_{k+1} \quad (2)$$

The auxiliary vector, g_{k+1} , is chosen to maximize, under fixed norm, the magnitude of the cross-correlation between its output, $g_{k+1}^H r$, and the previous filter output, $w_k^H r$, and is given by

$$g_{k+1} = R w_k - \frac{w_{R-MF}^H R w_k}{\|w_{R-MF}\|^2} w_{R-MF} \quad (3)$$

where R is in the input autocorrelation matrix, $R = E\{rr^H\}$. The scalar μ_{k+1} is selected such that it minimizes the output variance of the filter w_{k+1} or equivalently minimizes the mean square (MS) error between $w_k^H r$ and $\mu_{k+1}^* g_{k+1}^H r$. The MS-optimum μ_{k+1} is:

$$\mu_{k+1} = \frac{g_{k+1}^H R w_k}{g_{k+1}^H R g_{k+1}} \quad (4)$$

The AV filter recursion is completely defined by equations (1)-(4). The initial filter used for the AV technique, the RAKE filter, is estimated by using a certain number of pilot bits. These are obtained from the input bits that are assumed to be known at the receiver. The AV filter is used to despread the data stream. The performance of the AV was compared to the RAKE and minimum variance distortionless response (MVDR) filters.

A deinterleaver is used to restore symbols back to their original codewords. Now that any error bursts are randomized, the decoder can perform decoding as if the received data has gone through a random error channel instead of a burst error channel.

The Viterbi algorithm is most commonly used to decode convolutional codes. It is a maximum-likelihood sequence estimation procedure. Hard-decision decoding is used here. In hard decoding, the channel output is used to produce a maximum-likelihood decision on the channel input which is then used for the final source string decision making under the proper distance metric.

Results

The tabulated results obtained by this study are shown below. Since the resulting video quality of all the decoded frames is comparable, the standard by which the output is deemed superior is the number of frames that can be decoded. In all cases, the AV filter outperforms both the RAKE and MVDR filters. If there are multiple sets that produce fully decoded video clips, the best set is determined by the lowest percentage of the packet size used as pilot bits. The shaded row of each table represents the set of parameters that produce the best end result.

The packet size corresponds to the size of the packet before channel decoding. The packet size for channel decoding must remain at the original 410. In all cases, the six interfering users are cyclically shifted to create an asynchronous environment that occurs in real life.

Tables 1 and 2 give the results for the cases where there is no Doppler. Table 1 corresponds to the simulations run with the video clip at the highest data rate. With the original packet size of 410, 13% of the packet size must be known in order to obtain the full video using the AV filter. For the same packet size and number of pilot bits, RAKE and MVDR filters can not even produce decodable video.

FILTER	PACKET SIZE	PILOTS BITS	BER	# OF FRAMES DECODED
AV	410	13%	3.10E-06	ALL (All frames decoded)
RAKE	410	13%	5.90E-02	0 (Not decodable at all)
MVDR	410	13%	5.77E-04	0
AV	410	12%	1.40E-05	54
RAKE	410	12%	6.33E-02	0
MVDR	410	12%	4.20E-03	0
AV	410	11%	3.57E-05	27
AV	410	10%	4.03E-05	14
AV	410	9.75%	6.82E-05	0

Table 1: Results for 120kbps clip (142 frames)
No Doppler for all users

In Table 2, the performance is shown to improve if a larger packet size is used. Even with just 5% of the packet size used as pilot bits, the AV technique results in a 0 BER. The packet size may be further increased past 2008, but the computational complexity increases with increasing packet size.

FILTER	PACKET SIZE	PILOT BITS	BER	# OF FRAMES DECODED
AV	2008	10%	0	ALL
MVDR	2008	10%	0	ALL
AV	2008	5%	0	ALL
MVDR	2008	5%	2.54e-05	ALL

Table 2: Results for 40kbps clip (300 frames)
No Doppler for all users

The results for only the user of interest experiencing Doppler shifts of 4 Hz, 40 Hz, and 200 Hz while the interfering users experience no Doppler are given in Tables 3, 4, and 5, respectively. The packet sizes must be decreased with increasing Doppler frequencies to achieve adequate results. In Table 3, the AV filter gives fully decodable video with only 7.5% of the packet size used as pilot bits while MVDR requires 10% for pilot bits.

FILTER	PACKET SIZE	PILOT BITS	BER	# OF FRAMES DECODED
AV	810	10%	0	ALL
MVDR	810	10%	3.24E-05	ALL
AV	810	7.5%	0	ALL
MVDR	810	7.5%	1.09E-04	0
AV	810	5%	1.43E-04	0
MVDR	810	5%	2.2E-03	0

Table 3: Results for 40 kbps clip (300 frames)
4 Hz Doppler for User of Interest; No Doppler for Interferers

In Table 4, the AV method achieves fully decodable video with just 10% of a packet size utilized for pilot bits. The MVDR filter can not produce any decodable video even with 30% used for pilot bits. In fact, it requires the knowledge of half of the packet in order to result in video that is fully decodable which is not realistic.

FILTER	PACKET SIZE	PILOT BITS	BER	# OF FRAMES DECODED
MVDR	610	50%	0	ALL
MVDR	610	30%	7.63E-05	0
MVDR	610	10%	0.0015	0
AV	610	10%	0	ALL
AV	610	7.5%	6.01E-05	134
AV	610	5%	3.40E-04	0

Table 4: Results for 40 kbps clip (300 frames) with 40 Hz Doppler
40 Hz Doppler for User of Interest; No Doppler for Interferers

Even with a Doppler frequency of 200 Hz, the AV technique still achieved fully decodable video with just 12.5 % of the packet size used as pilot bits, as seen in Table 5. MVDR could not produce decodable video even with 50% of the packet size, and so it was omitted.

FILTER	PACKET SIZE	PILOT BITS	BER	# OF FRAMES DECODED
AV	250	12.5%	4.62E-06	ALL
AV	250	10%	2.48E-04	0

Table 5: Results for 40 kbps clip (300 frames)
200 Hz Doppler for User of Interest; No Doppler for Interferers

The situation where ALL users experience the same Doppler frequencies of 4 Hz, 40 Hz, and 200 Hz are shown in Tables 6, 7, and 8, respectively. There are three cases that produce fully decodable video using AV filters. However, using 30% for a packet size of 266 or even 20% for a packet size of 1000 for pilot bits is not a reasonable condition. Hence using 12.5% for a packet size of 800 is regarded as the best choice. Neither RAKE nor MVDR is able to generate any decodable video.

FILTER	PACKET SIZE	PILOT BITS	BER	# OF FRAMES DECODED
AV	266	30%	2.15E-05	ALL
RAKE	266	30%	0.0497	0
MVDR	266	30%	9.48E-04	0
AV	600	25%	5.41E-05	114
RAKE	600	25%	0.0448	0
MVDR	600	25%	1.76E-04	0
AV	800	12.5%	4.31E-05	ALL
RAKE	800	12.5%	0.05	0
MVDR	800	12.5%	3.90E-05	0
AV	800	10%	7.39E-05	70
RAKE	800	10%	0.0571	0
MVDR	800	10%	1.20E-03	0
AV	1000	20%	7.18E-06	ALL
RAKE	1000	20%	0.0572	0
MVDR	1000	20%	6.74E-05	70

Table 6: Results for 40 kbps clip (142 frames)
4 Hz Doppler for all users

An increase to 15% of a packet used for pilot bits is required when the Doppler is raised to 40 Hz, as shown in Table 7. MVDR is omitted since it could not produce any video that could be decoded even with an unreasonably large number of pilot bits.

FILTER	PACKET SIZE	PILOT BITS	BER	# OF FRAMES DECODED
AV	410	20%	5.85E-04	0
AV	600	15%	3.59E-05	ALL
AV	600	20%	5.31E-04	0
AV	800	15%	1.40E-03	0

Table 7: Results for 40 kbps clip (142 frames)
40 Hz Doppler for all users

Finally, for the last case where all users are experiencing the same Doppler frequency of 200 Hz, the AV technique fails as well. Only 66 frames are able to be decoded even with the knowledge of half the packet and a packet size as small as 100.

FILTER	PACKET SIZE	PILOT BITS	BER	# OF FRAMES DECODED
AV	100	50%	6.74E-05	66
AV	100	30%	9.59E-04	0

Table 8: Results for 40 kbps clip (142 frames)
200 Hz Doppler for all users

C Implementation of the Developed Algorithms

We have collaborated with AFRL personnel in order to convert the Matlab code we used for the above results to C++. Thus, the following modules are being implemented in C++:

- AV receiver
- MVDR receiver
- RAKE-MF receiver
- Deinterleaver
- Channel decoder

The C++ implementation is expected to eventually lead to an efficient hardware implementation of the proposed algorithms.

Publications

This research has led to the following publication:

M. Bansal, E. S. Pynadath, L. P. Kondi, J. D. Matyjas, M. J. Medley, S. P. Reichhart, K. A. Turck, “Interference Resistant Scalable Video Transmission over DS-CDMA Channels”, in Proc. MILCOM’05, Atlantic City, NJ, to appear.

This publication is attached.

Appendix A

INTERFERENCE RESISTANT SCALABLE VIDEO TRANSMISSION OVER DS-CDMA CHANNELS

Abstract ID#: 1374

Manu Bansal, Elizabeth S. Pynadath, Lisimachos P. Kondi
Department of Electrical Engineering,
University at Buffalo, SUNY,
Buffalo, NY 14260.
Email:{mbansal,pynadath,lkondi}@eng.buffalo.edu

and

John D. Matyjas, Michael J. Medley, Stephen P. Reichhart,
Kurt A. Turck
Air Force Research Laboratory/IFGC,
525 Brooks Road,
Rome, NY 13441.
Email:{john.matyjas,medley,reichhart,kurt.turck}@rl.af.mil

ABSTRACT

In this work, we demonstrate the interference mitigation capabilities of the auxiliary vector (AV) receiver for scalable video transmission over direct-sequence code division multiple access (DS-CDMA) systems using a hardware testbed. The proposed receiver design is also compared to the conventional RAKE matched-filter (RAKE-MF) and minimum variance distortionless response (MVDR) receivers. The DS-CDMA video data stream is transmitted over an RF channel under “real world” Rayleigh-faded multipath channel conditions, emulating open and/or urban battlefield environments. The state-of-the-art Agilent E4438C Vector Signal Generator and Baseband Studio Fader is used to provide a configurable “real time” RF channel. In this work, the “foreman” video sequence is source encoded using an MPEG-4 compatible video codec and channel-coded using rate-compatible punctured convolutional (RCPC) codes. After spreading and modulating, the resultant bitstream is transmitted over a user-defined Agilent wireless channel emulation. Upon chip-matched filtering and sampling at the chip-rate on a hardware testbed, the received data are despread/demodulated using the AV, RAKE-MF and MVDR receivers and, subsequently, channel and source decoded. The resultant video clips exemplify that the AV receiver outperforms the MVDR and the RAKE-MF receiver counterparts under a wide range of rates and channel conditions.

1 INTRODUCTION

In the last few years, there has been a significant amount of research work done in the area of multimedia communication over direct-sequence code division multiple access (DS-CDMA) channels. In this paper we consider the scalable video transmission over “real world” DS-CDMA multipath fading channels in a multiuser environment and demonstrate the interference suppression capabilities of the auxiliary-vector (AV) receiver [1] for such systems. The choice of the AV receiver was dictated by realistic channel fading rates that limit the data record available for receiver adaptation and redesign. Under small sample support adaptation, the AV filter short-data-record estimators have been shown to exhibit superior bit-error-rate performance in comparison with least-mean-squares (LMS), recursive-least-squares (RLS), sample-matrix-inversion (SMI), diagonally loaded SMI, or multistage nested Wiener filter implementations [1], [2], [3].

In [4], video transmission from one transmitter to one receiver using binary phase-shift-keying (BPSK) modulation was analyzed. The channel behavior was modeled as non-frequency-selective Rayleigh fading. In [5], video transmission over a DS-CDMA link was considered. A frequency-selective (multipath) Rayleigh fading channel model was used. At the receiver, an adaptive antenna array auxiliary-vector (AV) linear filter that provides space-time RAKE-type processing (thus, taking advantage of the multipath characteristics of the channel) and multiple-access in-

interference suppression was employed [1]. The trade-offs of source coding, channel coding and spreading for image transmission in CDMA systems were considered in [6]. In [7] and [8], video transmission via a single-rate CDMA channel was compared against transmission via a combination of multi-code multirate CDMA and variable sequence length multirate CDMA under frequency selective Rayleigh fading.

In the video transmission system proposed here, the scalable video bit stream is first channel encoded using a particular channel coding rate. The channel coded information is then spread using a spreading code and carrier modulated for transmission over the wireless channel. The transmission over an RF channel with Rayleigh-faded multipath channel conditions, emulating open and/or urban battlefield environments, is done by means of the Agilent E4438C Vector Signal Generator and Baseband Studio Fader. At the receiver the information is demodulated and despread. These despread data are then decoded using a channel decoder and the video sequence is finally reconstructed using the source decoder.

The rest of this paper is organized as follows. In section 2, we describe the elements of the proposed video transmission system, i.e., scalable video coding (section 2.1), channel encoding (section 2.2), channel modeling (section 2.3), the received signal (section 2.4), and auxiliary-vector (AV) filtering (section 2.5). Experimental results are presented in section 3 and conclusions are drawn in section 4.

2 VIDEO TRANSMISSION SYSTEM

2.1 Scalable Video Coding

A scalable video encoder produces a bit stream that consists of embedded subsets. Each embedded subset can be decoded and produce a video sequence of a certain quality. Thus, a single compression operation can produce bit streams with different rates and reconstructed quality. A subset of the original bit stream can be initially transmitted to provide a base layer quality with extra layers subsequently transmitted as enhancement layers.

In this work, an MPEG-4 compatible video source codec is used for scalable video coding. Scalability is obtained in terms of SNR where enhancement in quality translates in an increase in the SNR of the reconstructed video sequence [9]. Also, error-resilience tools of MPEG-4 such as resynchronization markers, data partitioning and reversible variable length codes

(RVLC) were enabled.

2.2 Channel Coding

Rate-Compatible Punctured Convolutional (RCPC) codes [10] are used to provide the unequal error protection (UEP) to the bitstream. The rate of a convolutional code is defined as k/n where k is the number of input bits and n is the number of output bits. For variable rate coding, a higher rate code can be obtained by puncturing the output of a “mother” code of rate $1/n$. For rate compatibility, higher rate codes are chosen to be subsets of lower rate codes. Decoding of the convolutional codes is carried out by the Viterbi algorithm which is a maximum-likelihood sequence estimation (MLSE) procedure.

2.3 Channel Modeling

The Agilent E4438C RF Arbitrary Waveform Generator and the N115A Baseband Studio fader hardware/software package were used to create a “real time” RF Rayleigh fading channel with multipath and Doppler frequencies. The E4438C RF Waveform Generator converts baseband I/Q data to RF. The N115A Baseband Studio obtains non-faded I/Q data from the E4438C arbitrary waveform generator and based on operator selected fading parameters, it computes and inserts, in “real time”, faded I/Q data back into the waveform generator. This setup is used to emulate open and/or urban battlefield environments with up to 48 paths and various Doppler frequencies.

2.4 Received Signal

We model the baseband received signal as the aggregate of the received multipath spread-spectrum (SS) signal of interest with signature code \mathbf{S}_o of length L (if T is the symbol period and T_c is the chip period then $L = T/T_c$), $K - 1$ received DS-SS interferers with unknown signatures \mathbf{S}_k , $k = 1, \dots, K - 1$, and white Gaussian noise. For notational simplicity and without the loss of generality, we choose a chip-synchronous signal set-up. We assume that the multipath spread is of the order of a few chip intervals, P , and since the signal is bandlimited to $B = 1/2T_c$ the lowpass channel can be represented as a tapped delay line with $P + 1$ taps spaced at chip intervals T_c . After conventional chip-matched filtering and sampling at the chip rate over a multipath-extended symbol interval of $L + P$ chips, the $L + P$ data samples from the antenna element are organized in the form of a vector \mathbf{r} given by

$\mathbf{r} = \sum_{k=0}^{K-1} \sum_{p=0}^P c_{k,p} \sqrt{E_k} (b_k \mathbf{S}_{k,p} + b_k^- \mathbf{S}_{k,p}^- + b_k^+ \mathbf{S}_{k,p}^+) + \mathbf{n}$, where, with respect to the k th SS signal, E_k is the transmitted energy, b_k , b_k^- , b_k^+ are the present, the previous and the following transmitted bits, respectively, and $\{c_{k,p}\}$ are the coefficients of the Rayleigh fading multipath channel emulated using the Agilent E4438C Vector Signal Generator and Baseband Studio Fader. $\mathbf{S}_{k,p}$ represents the zero-padded by P , p -cyclic-shifted version of the signature of the k th SS signal \mathbf{S}_k , $\mathbf{S}_{k,p}^-$ is the 0-filled $(L-p)$ -left-shifted version of $\mathbf{S}_{k,0}$ and $\mathbf{S}_{k,p}^+$ is the 0-filled $(L-p)$ -right-shifted version of $\mathbf{S}_{k,0}$. Finally, \mathbf{n} is the additive complex Gaussian noise.

For conceptual and notational simplicity we may rewrite the received data equation as follows: $\mathbf{r} = \sqrt{E_0} b_0 \mathbf{w}_{R-MF} + \mathbf{I} + \mathbf{n}$, where $\mathbf{w}_{R-MF} = E_{b_0} \{\mathbf{r} b_0\} = \sum_{p=0}^P c_{0,p} \mathbf{S}_{0,p}$ is the effective channel-processed signature (RAKE Matched-Filter) of the SS signal of interest (signal-0) and \mathbf{I} identifies comprehensively both the Inter-Symbol and the SS interference present in \mathbf{r} . In this work, as the fading coefficients are not known, $E_{b_0} \{\cdot\}$, the statistical expectation with respect to b_0 is used to estimate \mathbf{w}_{R-MF} . This is done by using pilot bits (of the SS signal of interest) which are assumed to be available at the receiver error-free.

2.5 Auxiliary-Vector Filtering

After carrier demodulation, chip-matched filtering, and chip-rate sampling, auxiliary-vector (AV) filtering [1] provides multiple-access-interference (MAI) suppressing despreading. The AV receiver was chosen based on the realistic channel fading rates that limit the data record available for receiver adaptation. Under small sample support adaptation, AV filter short-data-record estimators have been shown to exhibit superior bit error rate (BER) performance in comparison to least mean squares (LMS), recursive least squares (RLS), sample matrix inversion (SMI), diagonally-loaded SMI, or multistage nested Wiener filter implementations. The AV algorithm generates a sequence of AV filters making use of two basic principles: (i) The maximum magnitude cross-correlation criterion for the evaluation of the auxiliary vectors (ii) The conditional mean-squared optimization criterion for the evaluation of the scalar AV weights. This algorithm is more clearly explained below.

The AV algorithm generates an infinite sequence of filters $\{\mathbf{w}_k\}_{k=0}^\infty$. The sequence is initialized at the

RAKE filter

$$\mathbf{w}_0 = \frac{\mathbf{w}_{R-MF}}{\|\mathbf{w}_{R-MF}\|^2}, \quad (1)$$

which is normalized to satisfy $\mathbf{w}_0^H \mathbf{w}_{R-MF} = 1$. At each step $k+1$ of the algorithm, $k = 0, 1, 2, \dots$, we incorporate in \mathbf{w}_k an “auxiliary” vector component \mathbf{g}_{k+1} that is orthogonal to \mathbf{w}_{R-MF} and weighted by a scalar μ_{k+1} and we form the next filter in the sequence,

$$\mathbf{w}_{k+1} = \mathbf{w}_k - \mu_{k+1} \mathbf{g}_{k+1}. \quad (2)$$

The auxiliary vector \mathbf{g}_{k+1} is chosen to maximize, under fixed norm, the magnitude of the cross-correlation between its output, $\mathbf{g}_{k+1}^H \mathbf{r}$, and the previous filter output, $\mathbf{w}_k^H \mathbf{r}$, and is given by

$$\mathbf{g}_{k+1} = \mathbf{R} \mathbf{w}_k - \frac{\mathbf{w}_k^H \mathbf{R} \mathbf{w}_k}{\|\mathbf{w}_{R-MF}\|^2} \mathbf{w}_{R-MF} \quad (3)$$

where \mathbf{R} is the input autocorrelation matrix, $\mathbf{R} = E\{\mathbf{r} \mathbf{r}^H\}$. The scalar μ_{k+1} is selected such that it minimizes the output variance of the filter \mathbf{w}_{k+1} or equivalently minimizes the mean-square (MS) error between $\mathbf{w}_k^H \mathbf{r}$ and $\mu_{k+1}^* \mathbf{g}_{k+1}^H \mathbf{r}$. The MS-optimum μ_{k+1} is

$$\mu_{k+1} = \frac{\mathbf{g}_{k+1}^H \mathbf{R} \mathbf{w}_k}{\mathbf{g}_{k+1}^H \mathbf{R} \mathbf{g}_{k+1}}. \quad (4)$$

The AV filter recursion is completely defined by (1)-(4). Theoretical analysis of the AV algorithm was pursued in [1]. The results are summarized below in the form of a theorem.

Theorem 1: Let \mathbf{R} be a Hermitian positive definite matrix. Consider the iterative algorithm of eqs. (1)-(4).

(i) Successive auxiliary vectors generated through (2)-(4) are orthogonal: $\mathbf{g}_k^H \mathbf{g}_{k+1} = 0$, $k = 1, 2, 3, \dots$ (however, in general $\mathbf{g}_k^H \mathbf{g}_j \neq 0$ for $|k-j| \neq 1$).

(ii) The generated sequence of auxiliary-vector weights $\{\mu_k\}$, $k = 1, 2, \dots$, is real-valued, positive, and bounded: $0 < \frac{1}{\lambda_{\max}} \leq \mu_k \leq \frac{1}{\lambda_{\min}}$, $k = 1, 2, \dots$, where λ_{\max} and λ_{\min} are the maximum and minimum, correspondingly, eigenvalues of \mathbf{R} .

(iii) The sequence of auxiliary vectors $\{\mathbf{g}_k\}$, $k = 1, 2, \dots$, converges to the $\mathbf{0}$ vector: $\lim_{n \rightarrow \infty} \mathbf{g}_n = \mathbf{0}$.

(iv) The sequence of auxiliary-vector filters $\{\mathbf{w}_k\}$, $k = 1, 2, \dots$, converges to the minimum-variance-distortionless-response (MVDR) filter: $\lim_{k \rightarrow \infty} \mathbf{w}_k =$

$$\frac{\mathbf{R}^{-1} \mathbf{w}_{R-MF}}{\mathbf{w}_{R-MF}^H \mathbf{R}^{-1} \mathbf{w}_{R-MF}}. \quad \square$$

If \mathbf{R} is unknown (as in practice) and is sample-average estimated from a packet data record of D points, $\hat{\mathbf{R}}(D) = \frac{1}{D} \sum_{d=1}^D \mathbf{r}_d \mathbf{r}_d^H$, then Theorem 1 shows that

$$\hat{\mathbf{w}}_k(D) \xrightarrow{k \rightarrow \infty} \hat{\mathbf{w}}_\infty(D) = \frac{\left[\hat{\mathbf{R}}(D)\right]^{-1} \mathbf{w}_{\text{R-MF}}}{\mathbf{w}_{\text{R-MF}}^H \left[\hat{\mathbf{R}}(D)\right]^{-1} \mathbf{w}_{\text{R-MF}}} \quad (5)$$

where $\hat{\mathbf{w}}_\infty(D)$ is the widely used MVDR filter estimator known as the sample-matrix-inversion (SMI) filter [11]. The output sequence begins from $\hat{\mathbf{w}}_0(D) = \frac{\mathbf{w}_{\text{R-MF}}}{\|\mathbf{w}_{\text{R-MF}}\|^2}$, which is a θ -variance, fixed-valued, estimator that may be severely biased ($\hat{\mathbf{w}}_0(D) = \frac{\mathbf{w}_{\text{R-MF}}}{\|\mathbf{w}_{\text{R-MF}}\|^2} \neq \mathbf{w}_{\text{MVDR}}$) unless $\mathbf{R} = \sigma^2 \mathbf{I}$ for some $\sigma > 0$. In the latter trivial case, $\hat{\mathbf{w}}_0(D)$ is already the perfect MVDR filter. Otherwise, the next filter estimator in the sequence, $\hat{\mathbf{w}}_1(D)$, has a significantly reduced bias due to the optimization procedure employed at the expense of non-zero estimator (co-)variance. As we move up in the sequence of filter estimators $\hat{\mathbf{w}}_k(D)$, $k = 0, 1, 2, \dots$, the bias decreases rapidly to zero while the variance rises slowly to the SMI ($\hat{\mathbf{w}}_\infty(D)$) levels (cf. (5)).

An adaptive data-dependent procedure for the selection of the most appropriate member of the AV filter estimator sequence $\{\hat{\mathbf{w}}_k(D)\}$ for a given data record of size D is presented in [12]. The procedure selects the estimator $\hat{\mathbf{w}}_k$ from the generated sequence of AV filter estimators that exhibits maximum \mathcal{J} -divergence between the filter output conditional distributions given that +1 or -1 is transmitted. Under a Gaussian approximation on the conditional filter output distribution, it was shown in [12] that the \mathcal{J} -divergence of the filter estimator with k auxiliary vectors is

$$J(k) = \frac{4 E^2 \{b_0 \text{Re} [\hat{\mathbf{w}}_k^H(D) \mathbf{r}]\}}{\text{Var} \{b_0 \text{Re} [\hat{\mathbf{w}}_k^H(D) \mathbf{r}]\}}. \quad (6)$$

To estimate the \mathcal{J} -divergence $J(k)$ from the data packet of size D , the transmitted information bits b_0 are required to be known. We can obtain a blind *approximate* version of $J(k)$ by substituting the information bit b_0 in (6) by the detected bit $\hat{b}_0 = \text{sgn} [\text{Re} \{\hat{\mathbf{w}}_k^H(D) \mathbf{r}\}]$ (output of the sign detector that follows the linear filter). In particular, using \hat{b}_0 in place of b_0 in (6) we obtain the following \mathcal{J} -divergence

expression:

$$\begin{aligned} J_B(k) &= \frac{4 E^2 \{\hat{b}_0 \text{Re} [\hat{\mathbf{w}}_k^H(D) \mathbf{r}]\}}{\text{Var} \{\hat{b}_0 \text{Re} [\hat{\mathbf{w}}_k^H(D) \mathbf{r}]\}} \\ &= \frac{4 E^2 \{|\text{Re} [\hat{\mathbf{w}}_k^H(D) \mathbf{r}]\}|}{\text{Var} \{|\text{Re} [\hat{\mathbf{w}}_k^H(D) \mathbf{r}]\}|} \end{aligned} \quad (7)$$

where the subscript “B” identifies the blind version of the \mathcal{J} -divergence function. To estimate $J_B(k)$ from the data packet of size D , we substitute the statistical expectations in (7) by sample averages. The following criterion summarizes the corresponding AV filter estimator selection rule.

Criterion 1: For a given data record of size D , the unsupervised (blind) \mathcal{J} -divergence AV filter estimator selection rule chooses the estimator $\hat{\mathbf{w}}_k(D)$ with k auxiliary vectors where

$$\begin{aligned} k &= \arg \max_k \left\{ \hat{J}_B(k) \right\} = \\ &= \arg \max_k \left\{ \frac{4 \left[\frac{1}{D} \sum_{d=1}^D |\text{Re} [\hat{\mathbf{w}}_k^H(D) \mathbf{r}_d]| \right]^2}{\left[\frac{1}{D} \sum_{d=1}^D |\text{Re} [\hat{\mathbf{w}}_k^H(D) \mathbf{r}_d]| \right]^2 + \left[\frac{1}{D} \sum_{d=1}^D |\text{Re} [\hat{\mathbf{w}}_k^H(D) \mathbf{r}_d]| \right]^2} \right\} \end{aligned} \quad (8)$$

Criterion 1 completes the design of the AV filter estimator.

3 EXPERIMENTAL RESULTS

In this section, we present the experimental results for the setup described above. We assume $K = 7$ users that employ direct-sequence SS signaling (the user-of-interest and six interferers). The SNR for all the users is fixed at 8 dB (all SNR values reported in this paper refer to the SNR per chip). An MPEG-4 compatible source codec was used to encode a “foreman” video sequence with two different source rates of 40 and 120 kbps. Rate-compatible punctured codes (RCPC) were used for channel coding by using a “mother” code rate of 1/4 and puncturing it down to the code rate of 1/2. Walsh-Hadamard codes of length $L = 16$ were used as spreading codes. The transmission over a Rayleigh fading channel was simulated using the Agilent RF Waveform Generator and Baseband Studio Fader with all CDMA signals $k = 0, 1, \dots, K - 1$ experiencing $P = 3$ resolvable multipaths and various Doppler values equal to 0, 4, 40 and 200 Hz. Three different receivers were assumed: The RAKE matched-filter (RAKE-MF), the conventional sample-matrix-inversion minimum-variance-distortionless-response (SMI-MVDR) filter,

TABLE I

SIMULATION RESULTS FOR 150 FRAME VIDEO SEQUENCE
ENCODED AT 120 KBPS WITH ZERO DOPPLER.

Filter used	Packet size (bits)	Pilot bits	BER	no. of dec. frames
AV	410	12%	1.4E-05	54
RAKE-MF	410	12%	6.33E-02	0
MVDR	410	12%	4.20E-03	0
AV	410	13%	3.1E-06	150 (ALL)
RAKE-MF	410	13%	5.9E-02	0
MVDR	410	13%	5.77E-04	0

and the auxiliary vector (AV) filter (based on criterion 1, the best AV was selected out of 11 AV's that were produced). The channel decoding was performed using Viterbi algorithm.

The tabulated results obtained by this study are shown next. Since the resulting video quality of all the decoded frames is comparable, the standard by which the output is deemed superior is the number of frames that can be decoded. If there are multiple sets that produce fully-decodable video clips (all the frames), the best set is determined by the lowest number of pilot bits used (given in terms of percentage of the packet size). The packet size in the tables below refer to the size of the packet after channel encoding. For all the cases, the bit error rate (BER) is calculated after channel decoding at the receiver. Table I gives the results for the cases where there is zero Doppler. Table I corresponds to the simulations run with the video sequence (150 frames) encoded at the rate of 120 kbps. Using 13%/12% of the packet size (410 bits) as pilot bits, AV filter recovered the total/partial video sequence, respectively. For the same packet size and number of pilot bits, the RAKE-MF and the MVDR filter can not even produce decodable video.

The results for only the user-of-interest experiencing Doppler shifts of 4 Hz while the interfering users experience zero Doppler are given in Table II. The packet sizes were adjusted according to the Doppler frequency to maintain the assumption of constant fading over each packet. As shown in Table II, the AV filter gives a fully decodable video with only 7.5% of the packet size used as pilot bits while the MVDR filter requires higher percentage for pilot bits. Using the RAKE-MF filter, no frames were recovered with a practical number of pilot bits.

Table III shows the results for the situation where all

TABLE II

SIMULATION RESULTS FOR 300 FRAME VIDEO SEQUENCE
ENCODED AT 40 KBPS WITH 4 HZ DOPPLER.

Filter used	Packet size	Pilot bits	BER	no. of dec. frames
AV	810	7.5%	0	300
RAKE-MF	810	7.5%	4.68E-02	0
MVDR	810	7.5%	1.09E-04	0
AV	810	10%	0	300
RAKE-MF	810	10%	2.54E-02	0
MVDR	810	10%	3.24E-05	300

TABLE III

SIMULATION RESULTS FOR 150 FRAME VIDEO SEQUENCE
ENCODED AT 40 KBPS WITH 4 HZ DOPPLER.

Filter used	Packet size	Pilot bits	BER	no. of dec. frames
AV	800	10%	7.39E-05	70
RAKE-MF	800	10%	5.71E-02	0
MVDR	800	10%	1.20E-03	0
AV	800	12.5%	4.31E-05	150
RAKE-MF	800	12.5%	5E-02	0
MVDR	800	12.5%	3.90E-05	0
AV	1000	20%	7.18E-06	150
RAKE-MF	1000	20%	5.72E-02	0
MVDR	1000	20%	6.74E-05	70
AV	600	25%	5.41E-05	150
RAKE-MF	600	25%	4.48E-02	0
MVDR	600	25%	1.76E-04	0
AV	266	30%	2.15E-05	150
RAKE-MF	266	30%	4.97E-02	0
MVDR	266	30%	9.48E-04	0

users experience the same Doppler frequencies of 4 Hz. There are three cases that produce fully-decodable video using AV filters. However, using 30% for a packet size of 266 or even 20% for a packet size of 1000 bits for pilot bits is not a practical condition. Hence using 12.5% for a packet size of 800 bits as pilot bits can be regarded as the best choice. Also, neither RAKE-MF filter nor MVDR filter recovered any part of the video sequence for any reasonable number of pilot bits.

Tables IV and V show the results for 40 and 200 Hz Doppler shifts. The values in these tables clearly show that the AV filter outperforms RAKE-MF and MVDR filters for all range of practical number of pilot bits used.

TABLE IV

SIMULATION RESULTS FOR 300 FRAME VIDEO SEQUENCE
ENCODED AT 40 KBPS WITH 40 HZ DOPPLER.

Filter used	Packet size	Pilot bits	BER	no. of dec. frames
AV	610	7.5%	6.01E-05	134
RAKE-MF	610	7.5%	5.8E-02	0
MVDR	610	7.5%	6.5E-03	0
AV	610	10%	0	300
RAKE-MF	610	10%	7.2E-02	0
MVDR	610	10%	1.5E-03	0

TABLE V

SIMULATION RESULTS FOR 300 FRAME VIDEO SEQUENCE
ENCODED AT 40 KBPS WITH 200 HZ DOPPLER.

Filter used	Packet size	Pilot bits	BER	no. of dec. frames
AV	250	12.5%	4.62E-06	300
RAKE-MF	250	12.5%	5.7E-02	0
MVDR	250	12.5%	2.48E-04	0

4 Conclusions

We demonstrated the effectiveness of using an auxiliary vector (AV) receiver for scalable video transmission over DS-CDMA systems with “real world” multipath channel conditions. The results clearly establish the interference mitigation capabilities of the AV receiver. The AV filter receiver was also shown to outperform the MVDR and the RAKE-MF receivers under a wide range of rates and channel conditions.

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